METHOD AND APPARATUS FOR TRANSMITTING SIGNALS IN A MULTI-ANTENNA MOBILE COMMUNICATIONS SYSTEM THAT COMPENSATES FOR CHANNEL VARIATIONS

Field of the Invention

5   The present invention relates to wireless communications, and more particularly, to transmitting signals from a transmitter comprising two or more antennas in a mobile telecommunications network.

Description of the Related Art

In mobile telecommunications systems, even though feeding back information about a channel to a transmitter would produce benefits by enabling the transmitter

10   to compensate for channel variations, reliable feedback about a channel is difficult to obtain due to the rapid variations over time that a channel experiences.

When a mobile user terminal moves at high speed, the combined effects of channel estimation, power control, quantisation and feedback delay, prevent

15   feedback schemes which are "closed loop", i.e. involving feedback from the receiver to the transmitter, from accurately tracking rapid channel fluctuations. As a result, due to the imperfect information about the channel at both the transmitter and receiver, performance tends to worsen as the speed of a moving mobile increases. On the other hand, performance of space-time block coding schemes, which are

20   "open-loop" in not relying on feedback from the receiver, tend to improve with increasing mobile speeds, becoming increasingly less sensitive to channel estimation imperfections with increasing speed.

It has been shown that if there were perfect channel feedback the optimal transmission strategy is transmission in a single direction specified by the feedback.

25   This is known as a beamforming strategy. Conversely, it has been shown that with no channel feedback the optimum strategy is to transmit equal power in orthogonal independent directions. This is known as a diversity strategy. However, in practice, feedback information is often neither perfect nor zero.

Imperfect feedback about channels has started to be considered in the known

30   art. In one known approach, a given transmission strategy is classified according to the rank of its input spatial covariance matrix. For instance, a beamforming strategy corresponds to a rank-one matrix while a covariance matrix with full rank indicates a diversity strategy.

Summary of the Invention

An embodiment of the present invention is a method of transmitting signals from a transmitter comprising two or more antennas in a mobile telecommunications network. The method involves determining channel state information, estimating the

5   reliability of that channel state information, and space time block encoding at least one data sequence. Before transmitting the data sequence, a linear transformation is applied to the data sequence so as to at least partially compensate for channel variations. The linear transformation is dependent upon channel state information and upon the estimated reliability of channel state information.

10   Other embodiments of the present invention also provide a corresponding transmitter, receiver, and network for mobile telecommunications.

The inventor considered how, in order to improve performance for a wide range of mobile speeds, channel state information could be used in conjunction with space-time codes. In consequence, a linear precoding function is provided that is

15   adapted with respect to both channel state information and channel state information reliability.

Embodiments of the invention can handle both extremes of channel knowledge, i.e. where there is perfect channel knowledge at the transmitter and where there is no channel knowledge at the receiver, and all cases in between these

20   two extremes. Between the two extremes, available channel information is taken into account considering channel state information reliability so as to improve the space-time codes performance. Performance can be improved in both networks involving closed loop feedback and those involving open loop feedback.

An embodiment of the invention is a Universal Mobile Telecommunications

25   System (UMTS) space-time coded system, where knowledge of the reliability of channel state information is used together with the channel state information itself in order to adapt the transmitter to varying channel state information and channel state information reliability conditions. Performance degradation due to limited channel state information reliability is thus combated and robustness achieved.

30   ## Brief Description of the Drawings

Embodiments of the present invention will now be described by way of example and with reference to the drawings, in which:

Figure 1 is a diagram illustrating a MIMO telecommunications system,

Figure 2 is a diagram illustrating a transmitter and receiver implementation in UMTS, and

Figure 3 is a diagram illustrating an alternative transmitter and receiver implementation in UMTS.

5   **Detailed Description**

For ease of understanding a more general description is presented followed by an explanation of implementation aspects in a mobile telecommunications network of Universal Mobile Telecommunications System (UMTS) type. It should be noted that the present invention has applications not only in UMTS, but also, by way

10   of example and without limitation, in communication systems such as code division multiple access (CDMA) and wideband code division multiple access (W-CDMA).

MIMO systems, for example for use in UMTS, typically involve space-time block encoding. An example of space-time block encoding scheme for two transmit antennas is presented in S.M. Alamouti, "A Simple Transmit Diversity Technique for

15   Wireless Communications", IEEE Journal on Selected Areas in Communication, Vol. 16, No. 8, pp. 1451-1458, October 1998 and is also described in United States Patent US 6,185,258. The encoding and transmission sequence for this scheme is as follows: at a first transmission time instant t_1 symbols x_1 and x_2 are transmitted from antennas 1 and 2 respectively and at the next transmission instant t_2 symbols -

20   x_2^* and x_1^* are transmitted from antennas 1 and 2 respectively, where * denotes complex conjugate. This transmission sequence Z can be represented in matrix form

as $\begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}$.

As further background, it is known that the performance of MIMO systems can sometimes be improved by precoding. Precoding means applying a linear

25   transformation to symbols. For example, a linear precoder for space time coded systems which takes account if channel state information is described in a paper by S. Zhou and G.B. Giannakis entitled "Optimal transmitter eigen-beamforming and space-time block coding based on channel mean feedback", IEEE Transactions on Signal Processing, Vol. 50, n°10, October 2002.

30   A linear precoder that exploits knowledge of channel state information is included in a space-time coded MIMO system so as to enhance performance. Let us consider a multiple-input multiple output (MIMO) telecommunications network

consisting of M transmit antennas and N receive antennas, as shown in Figure 1. The transmitter is, for example, a base station for mobile telecommunications.

5   For the analysis below, the network is considered to have one receive antenna (i.e. N=1); however, the approach presented can be readily generalised to any number of receive antennas. Without loss of generality, let us consider a channel which is non-selective in frequency. Each entry h_i of the Mx1 channel vector **h** represents the channel state, also known as the channel transfer function or channel response, between transmit antenna i and the receive antenna. Some

10   information on channel state is known at the transmitter and at the receiver.

As depicted in Figure 1, a space-time encoder maps the input data sequence $\mathbf{x}=(x_1, x_2, ..x_Q)$ to be transmitted into a QxM matrix **Z** of codewords, that are split on a set of M parallel sequences. These codewords are then transformed by a MxM linear transformation denoted **L** in order to adapt the code to the available channel

15   information. The resulting sequences, which are represented by the columns of a QxM matrix **C=ZL**, are sent from the M transmit antennas over Q time intervals. The received signal included in a Qx1 vector **y** is then a superposition of M transmitted sequences corrupted by an additive white Gaussian noise characterised by the Qx1 vector **n** and with covariance matrix equal to $\sigma^2 I_Q$:

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$$\mathbf{y} = \mathbf{Ch} + \mathbf{n} \qquad (1)$$

A block-fading model is assumed, i.e. that the channel remains constant over a number of symbol periods (spanning a space-time codeword) and then changes in

25   a statistically independent fashion for the following codeword period.

At the receiver the symbol-by-symbol detection is performed. The receiver is a maximal ratio combining (MRC) receiver, that is one which maximises the signal to noise ratio for a given precoder matrix **C**. The problem to be addressed is how to select an optimal precoder matrix **C**, regardless of whether perfect or imperfect

30   channel state information is available at the transmitter.

Channel state information (CSI) means channel estimates, which may be quantised due to the limited feedback bandwidth. The channel state information is partial in the sense that it may not be completely accurate and hence not be completely reliable.

Based on the (partial) channel state information (CSI) f(h) available at the receiver, the channel **h** conditional on that information (and so denoted the conditional channel) is now considered. For this analysis, **h** and **f** are assumed to follow a Gaussian distribution $N(0, \sigma_h^2 \mathbf{I})$.

5   Denoting the expected value of the conditional channel as

$$E\{\mathbf{h} \mid \mathbf{f}\} = \hat{\mathbf{h}} \quad (2),$$

the conditional channel can be expressed as a function of the partial CSI function **f** as:

10    $$\mathbf{h} \mid \mathbf{f}(\mathbf{h}) = \hat{\mathbf{h}} + \varepsilon \quad\quad (3)$$

where ε is the channel estimation error and ε has a variance of σ_ε^2.

The optimal precoder matrix **C** is computed based on minimisation of symbol-error-rate (SER). A closed form expression for SER as described in the Zhou and
15   Giannakis paper mentioned above is used, and the bound of SER known as the Chernoff bound will be used for this analysis. Background to the Chernoff bounds is provided in M.K. Simon, M.S. Alouini, "Digital Communication over Generalised Fading Channels: A Unified Approach to the Performance Analysis", John Wiley & Sons, 2000.

20   In order to compute the bound, the signal to noise ratio at the receiver must first be determined.

According to the above described signal model defined by Equation (1), the signal-to-noise-ratio γ at the MRC receiver can be written as:

25    $$\gamma = E\left\{\left|\mathbf{h}^H \mathbf{C}^H \mathbf{C} \mathbf{h} s\right|^2\right\} \Big/ E\left\{\mathbf{h}^H \mathbf{C}^H \mathbf{n} \mathbf{n}^H \mathbf{C} \mathbf{h}\right\} = \mathbf{h}^H \mathbf{C}^H \mathbf{C} \mathbf{h} \, E_s / \sigma^2 \quad (4)$$

where E is expectation, **h** is channel vector, S is the transmitted symbol, H denotes the Hermitian of the matrix, σ^2 is the noise variance, and E_S is the average symbol energy. Symbols are of an m-PSK constellation.

30   Based on the following eigen-decomposition:

$$\mathbf{C}^H \mathbf{C} = \mathbf{U}_C \Lambda_C \mathbf{U}_C^H \quad (5)$$

where **U** is the matrix of eigenvectors of $\mathbf{C}^H\mathbf{C}$, $\mathbf{U}_C$ is unitary and

$\Lambda_C = diag(\delta_1, \delta_2 ...., \delta_M)$ (in other words Λ_C is the diagonal matrix of the eigenvalues

of $\mathbf{C}^H\mathbf{C}$) with δ_i the i-th eigenvalue of $\mathbf{C}^H\mathbf{C}$, the signal-to-noise-ratio γ is a summation

of the signal to noise ratios (SNRs) for each sub-channel, (each sub-channel

corresponding to a transmit antenna) :

$$\gamma = \sum_{i=1}^{M} \gamma_i \qquad (6)$$

It was shown in the previously mentioned Zhou and Giannakis paper that the

average SNR on each sub-channel is:

$$\overline{\gamma}_i = \delta_i(1 + K_i)\sigma_\varepsilon^2 \, E_s / \sigma^2 \qquad (7)$$

where $K_i = \left\|\left[\mathbf{U}_C^{\,H}\rho\mathbf{f}\right]_i\right\|^2 / \sigma_\varepsilon^2$, where ρ denotes the correlation coefficient

between the actual channel, i.e. channel vector **h**, and the channel state information

(CSI) denoted **f** .(This is defined in Equation 15 below).

The Chernoff bound on SER has been proved in the previously mentioned

Simon and Alouini paper to be:

$$SER_{BOUND} = \alpha \prod_{i=1}^{M} \frac{1}{1 + \delta_i \xi} \exp\left(-\frac{K_i \delta_i \xi}{1 + \delta_i \xi}\right) \qquad (8)$$

where $\alpha = \dfrac{(m-1)}{m}$, $\xi = g_{m-PSK}\sigma_\varepsilon^2 E_s / \sigma^2$, with $g_{m-PSK} = \sin^2\left(\dfrac{\pi}{m}\right)$.

The optimal precoder is based on minimisation of a pairwise error probability

(union bound). Specifically, the optimal precoder matrix **C** is the solution to the

following optimisation problem, in order to minimise the symbol error rate:

$$\min_{\mathbf{C}} SER_{BOUND}$$
$$\text{s.t. trace}\{\mathbf{C}^H\mathbf{C}\} = 1 \qquad (9)$$

This means minimising with respect to **C** the SER$_{BOUND}$ subject to the summation of the diagonal elements of **C**^{H}**C** representing transmitted power being constrained as equal to one.

5   Let now the correlation matrix of the mean conditional channel

$$E\{\mathbf{h} \mid \mathbf{f}\} \, E\{\mathbf{h} \mid \mathbf{f}\}^H = \hat{\mathbf{h}}\hat{\mathbf{h}}^H = \mathbf{U}_h \mathbf{D}_h \mathbf{U}_h^H \qquad (10)$$

where $\mathbf{U}_h$ is the matrix of eigenvectors of the correlation matrix of the mean

10   conditional channel, $\mathbf{D}_h = diag\left(\left\|\hat{\mathbf{h}}\right\|^2, 0, .., 0\right)$ is the matrix of eigenvalues of the

correlation matrix of the mean conditional channel, and $\lambda = \left\|\hat{\mathbf{h}}\right\|^2$. $\left\|\hat{\mathbf{h}}\right\|$ is, of course,

the norm of $\hat{h}$.

It was proved in the previously mentioned Zhou and Giannakis paper that the optimal precoder **C** has

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$$\mathbf{U}_C = \mathbf{U}_h \qquad (11)$$

Combining the equations (5) and (11), the optimal precoder matrix

$$\mathbf{C} = \mathbf{ZL} = \mathbf{Z}\Lambda_C^{\frac{1}{2}}\mathbf{U}_h^H \qquad (12)$$

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(with the unitary matrix **Z** being the space-time encoder) is a generalised beamformer with multiple beams pointing to orthogonal directions along the eigenvectors of the correlation matrix of the mean conditional channel. The power loading along the eigenvectors is determined by the matrix Λ_C.

The power constraint in (9), equivalent to the average transmitted symbol

25   power being E$_s$, is used to identify the optimal $\Lambda_C = diag(\delta_1, \delta_2...., \delta_M)$:

$$\delta_2 = ... = \delta_M = \left[\frac{2a}{b + \sqrt{b^2 - 4ac}} - \frac{1}{\xi}\right]_+ \qquad (13)$$

$$\delta_1 = 1 - (M-1)\delta_2 \qquad (14)$$

where $a = (1 + M/\xi)^2$, $b = \lambda/(\xi\sigma_\varepsilon^2) + (1 + M/\xi)(2M - 1)$, $c = M(M - 1)$, and $(\cdot)_+$ stands for $max(\cdot, 0)$ i.e. that the expression in parentheses takes its actual computed value if positive else is set to zero if negative.

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Application to a Two antenna transmission system

The application of the proposed precoder matrix **C** to a 2x1 (2 transmit antennas, 1 receive antenna) system with Alamouti space-time block coding is now considered. In this case, as previously mentioned:

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$$\mathbf{Z} = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}.$$

The computation of the precoder **C** given in equation (12) is based on the assumptions that the noise follows a Gaussian distribution, that is $\mathbf{n} \sim N(0, \sigma^2 I)$ of

15   known covariance σ^2, and the conditional channel $\mathbf{h}| \; \mathbf{f(h)} \sim N(\rho\mathbf{f}, \sigma_\varepsilon^2 I)$, that is the conditional channel follows a Gaussian distribution with mean $\rho\mathbf{f}$ and covariance $\sigma_\varepsilon^2 I$ where, as previously mentioned **f** is the partial CSI information available at the receiver, and ρ the correlation coefficient between the actual channel h and the CSI information **f**. ρ represents the reliability of the channel state information (CSI) and is

20   such that :

$$E\{\mathbf{hf}^H\} = \rho\sigma_h^2\mathbf{I} \qquad (15)$$

and

25

$$\sigma_\varepsilon^2 = \sigma_h^2(1 - |\rho|^2) \qquad (16)$$

is the variance of the error. σ_h^2 is, of course, the variance of channel vector h.

Considering three cases dependent upon the value of ρ as follows:

30   1)When the CSI reliability ρ is such that $0 < \rho < 1$, the linear transformation **L** in Equation (12) in the case of a 2 transmit antenna system is written as:

$$\mathbf{L} = \begin{bmatrix} \sqrt{\delta_1} & 0 \\ 0 & \sqrt{\delta_2} \end{bmatrix} \begin{bmatrix} w_1^* & w_2^* \\ w_2 & -w_1 \end{bmatrix}$$

where δ_1, δ_2 are described in Equation (13) and $\mathbf{U}_h = \begin{bmatrix} w_1 & w_2^* \\ w_2 & -w_1^* \end{bmatrix}$ in Equation (10).

2) When the CSI reliability $\rho = 0$, the Chernoff bound on SER given in Equation (8) becomes:

$$SER_{BOUND} = \alpha \prod_{i=1}^{M} \frac{1}{1 + \delta_i \xi}$$

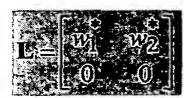and the solution to the minimisation problem (9) results in this case in:

$\delta_1 = \delta_2 = \cdots = \delta_M = \dfrac{1}{M}$. The linear transformation **L** in the case of a 2 transmit antenna system becomes:

$$\mathbf{L} = \frac{1}{\sqrt{2}} \begin{bmatrix} w_1^* & w_2^* \\ w_2 & -w_1 \end{bmatrix}$$

The transmission scheme is equivalent to 2x1 orthogonal space-time coding (Alamouti).

3) When the CSI reliability $\rho \rightarrow 1$, then, according to Equation (16), $\sigma_\varepsilon^2 \rightarrow 0$ and $\xi \rightarrow 0$. In this case $\delta_2 = \ldots = \delta_M = \left[\dfrac{2a}{b + \sqrt{b^2 - 4ac}} - \dfrac{1}{\xi} \right]_+ = 0$, as the expression in the brackets takes negative values. Only one eigenvalue is non-zero, $\delta_1 = 1$ and the linear transformation **L** in the case of a 2 transmit antenna system becomes:

$$\mathbf{L} = \begin{bmatrix} w_1^* & w_2^* \\ 0 & 0 \end{bmatrix}$$

The transmission scheme is then equivalent to a beamformer.

Computation of Channel State Information (CSI) reliability

The evaluation of channel state information (CSI) reliability depends on the nature of the CSI information. Under the assumption that the actual channel coefficients are fed back from the receiver to the transmitter after some delay $D(QT_c)$, i.e. the delay being a function of Q the codeword length and T_c the chip duration, then feedback vector **f**, i.e. channel state information, is taken to be

$$\mathbf{f}(\mathbf{h}(i)) = \mathbf{h}(i - D) \qquad (17)$$

In words, this means that the fedback CSI at time i is considered to be the actual channel at time i - D.

In this case the CSI reliability is the correlation coefficient ρ where

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$$E\{\mathbf{h}(i)\mathbf{h}(i-D)^H\} = \rho\sigma_h^2\mathbf{I} \qquad (18)$$

The delay D can be assumed equal to the feedback period, that is a timeslot. The computation of the CSI reliability ρ is based on a running average over

10   consecutive channel feedback samples. This averaging takes into account the previous feedback samples by introducing a forgetting factor α. The forgetting factor α aims to weight the contribution of each new channel estimate as compared to the past channel estimates, In mathematical terms, the reliability of CSI follows:

$$\rho(1) = 0$$

15            $$\rho(n) = (1 - \alpha)\mathbf{h}\mathbf{f}^* + \alpha\rho(n-1) \qquad (19)$$

The value of α is selected to be small when the CSI varies fast. This is so that the effect of past samples is smaller then. Conversely, when the CSI is more stable, more account is taken of past samples.

It should be noted that the channel state information (CSI) might be imperfect

20   due to other factors besides feedback delay. For example, it may be a quantised version of the channel or a mean over channel estimates over the timeslot, due to bandwidth limitations, as is the case in Universal Mobile Telecommunications System (UMTS) networks where, according to the relevant standard, a small number of feedback bits is used.

25

UMTS system architecture

Consider now application to a Universal Mobile Telecommunications System (UMTS) system with frequency division duplex (FDD) downlink. No assumption is made as to whether the channel is frequency selective or otherwise. In frequency

30   division duplex (FDD) uplink (i.e. from a mobile user terminal to a base station) and downlink (i.e. from a base station to a mobile user terminal) frequencies are different, so channel state information or equivalent information is fed back to the transmitter (base station).

As regards the transmission scheme and channel state information (CSI) reliability, CSI reliability takes into account channel estimates for previous feedback periods (e.g. timeslots) as outlined in Equations (17), (18) and (19). The CSI reliability is also affected by Doppler shift, feedback quantisation, and channel

5   estimation errors.

The computation of the linear transformation matrix L requires the CSI **f** and the CSI reliability ρ for the previous timeslot. Referring back to Figure 12, the matrix Z of codewords is known, Λ_C is determined from equation 13, and $U_h{}^H$ is determined from **f** and ρ and Equation 10 where $\hat{\mathbf{h}} = \rho\mathbf{f}$.

10   In the receiver, the linear transformation matrix **L** applied in transmission has to be estimated and given to the space time decoder for each tap of its Rake receiver. The receiver is similar to the one used with Space-Time Block Codes (and described in the Alamouti paper referred to above), except the linear transformation matrix L is taken into account. Indeed after despreading, the received signal on each

15   tap p is:

$$\mathbf{Y}^p = \begin{bmatrix} y_1^p \\ y_2^p \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \left(\sqrt{\delta_1}\,w_1 x_1 + \sqrt{\delta_2}\,w_2^* x_2\right)h_1^p + \left(\sqrt{\delta_1}\,w_2 x_1 - \sqrt{\delta_2}\,w_1^* x_2\right)h_2^p \\ \left(-\sqrt{\delta_1}\,w_1 x_2^* + \sqrt{\delta_2}\,w_2^* x_1^*\right)h_1^p - \left(\sqrt{\delta_1}\,w_2 x_2^* + \sqrt{\delta_2}\,w_1^* x_1^*\right)h_2^p \end{bmatrix} + \Sigma^p$$

(20)

20   where $\delta_1 + \delta_2 = 1, \delta_1, \delta_2 \geq 0$, i.e., $\delta_1 = \frac{1+\beta}{2}, \delta_2 = \frac{1-\beta}{2}$, and Σ^p is the noise matrix on tap p and subscript on signals y refers to received symbols at two time intervals. To recover transmit signals x_1 and x_2, the following operations are applied to the received signal $\mathbf{Y}^p$:

25
$$\begin{bmatrix} \hat{x}_1^p \\ \hat{x}_2^p \end{bmatrix} = \begin{bmatrix} \left(H_1^p\right)^* y_1^p + H_2^p \left(y_2^p\right)^* \\ \left(H_2^p\right)^* y_1^p - H_1^p \left(y_2^p\right)^* \end{bmatrix}$$
(21)

where

$$\begin{bmatrix} H_1^p \\ H_2^p \end{bmatrix} = \frac{1}{\sqrt{2}} \begin{bmatrix} \sqrt{\delta_1}\left(w_1 h_1^p + w_2 h_2^p\right) \\ \sqrt{\delta_2}\left(w_2^* h_1^p - w_1^* h_2^p\right) \end{bmatrix}$$
(22)

are the composite channels on tap p. The combiner block then combines the detected symbols on each tap.

UMTS system implementation

5   A UMTS transmitter 2 and receiver 4 are shown in Figure 2. The UMTS frequency division duplex (FDD) downlink transmission-reception scheme includes CSI (**f**) and CSI reliability (ρ) dependent linear precoding as explained above. The transmitter 2 thus has some knowledge (**f**, ρ) about the channel, which has been fedback. In a UMTS network operating FDD downlink (i.e. from base station to

10   mobile station), the CSI is obtained as feedback channel estimates 6 provided as bits sent by receiver 4 (i.e. the mobile station). The relevant modules are, at the transmitter, a linear precoder (**L**) 8, a processor 10 which receives the CSI and determines the CSI reliability, and L determinator 12. The relevant modules at the receiver are a processor 14 which determines the CSI reliability, an L determinator

15   16, and a space-time decoder 18.

At the transmitter 2, the linear precoder (**L**) 8 is applied to the space-time encoded symbols provided from a space-time block encoder 20 after channel coding, rate matching interleaving, and modulation (shown as functional clock 22) in known fashion. The linear precoder L coefficients are computed based on the CSI

20   and CSI reliability in the L determinator 12. The computation of CSI reliability (ρ) in the processor 10 is based on CSI i.e. channel estimates 6 fed back from the receiver 4. It is performed by averaging over time-sequential channel estimates (running average) using a forgetting factor. The forgetting factor aims to weight the contribution of present CSI as compared to the past CSI. This information is fed back

25   to the transmitter using a low-rate feedback link, as available in UMTS. In a UMTS uplink channel there is a number of bits available for communicating information to the transmitter about the received signal. The outputs of the linear precoder 8 are spread/scrambled 9 and subject to addition of Common Pilot Channel (CPICH) coding 11 bits before transmission.

30   At the receiver 4, received signals are used to provide CSI i.e. estimates (**f**) in a channel estimation block 26 so as to be used to compute the CSI reliability (ρ) in processor 14 (as at the transmitter). The signals are also despread 28 and applied to a space-time block decoder 24. At the receiver 4 the space-time block decoder (STD,24) has essentially the same structure as a conventional one (described in the

Alamouti paper and patent referred to above), but needs to consider instead of the channel estimates, the equivalent channel, defined as the linear transformation of the channel according to the coefficients of L, that is $\mathbf{H}_{eq} = \lfloor h_{eq,1} \quad h_{eq,2} \rfloor = \mathbf{HL}$. As

5 shown in Figure 2, the linear precoder L coefficients are estimated in the receiver by an L determinator 16. This has inputs from the channel estimation block 26 and the processor 14 which determines CSI reliability. The outputs of the space-time decoder 24 are provided to a combiner 30 and then channel decoded, inverse rate matched, deinterleaved and demodulated in known fashion (shown in Figure 2 as functional block 32).

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Alternative UMTS Implementation

An alternative implementation is now described, in which instead of the linear transformation matrix L being determined at the transmitter from CSI i.e. channel estimates provided by the receiver, the coefficients of linear transformation matrix L

15 are provided by the receiver.

In this alternative embodiment, which is shown in Figure 3, the transmitter 2' is given the coefficients of the precoder L by the receiver 4'. In this UMTS transmitter 2' and receiver 4' operating with frequency division duplex (FDD) downlink, these coefficients are feedback bits sent by the mobile station. The proposed UMTS

20 network is depicted in Figure 3, where the UMTS FDD downlink transmission-reception scheme includes CSI and CSI reliability dependent linear precoding as explained previously. The relevant module at the transmitter is a linear precoder (L) 8'. The relevant modules at the receiver are a processor 14' which determines the CSI reliability, an L determinator 16', and a space-time decoder 24'.

25 At the transmitter, the linear precoder (L) is applied to the space-time encoded symbols provided from the space-time block encoder 20' after channel coding, rate matching, interleaving, and modulation (shown as functional block 22') in known fashion. The outputs from the linear precoder 8' are spread/scrambled 9' and subject to the addition of Common Pilot Channel (CPICH) 11' bits before

30 transmission. The linear precoder L coefficients are provided by the receiver 4' as explained below and fed back over air to the transmitter.

At the receiver, the computation of CSI reliability (in a functional block denoted 14') is based on CSI i.e. channel estimates provided from the channel estimator block 26'. It is performed by averaging over time sequential channel

estimates (running average) using a forgetting factor. The forgetting factor aims to weight the contribution of the present CSI as compared to the past CSI ; the aim being to take account of slowly-varying CSI reliability. The linear precoder (L) coefficients are computed based on CSI and CSI reliability in the L determinator 16'.

5   At the receiver 4', received signals are used to provide channel estimates in a channel estimation block 26', and after despreading 28' are applied to a space-time decoder 24'. At the receiver, the space-time block decoder (STD) 24' has essentially t he same structure as the conventional one (described in the Alamouti paper and patent referred to above), but needs to consider instead of the channel estimates,

10   the equivalent channel, defined as the linear transformation of the channel according to the coefficients of L, that is $\mathbf{H}_{eq} = \begin{bmatrix} h_{eq,1} & h_{eq,2} \end{bmatrix} = \mathbf{HL}$. The outputs of the space-time decoder are provided to a combiner 30' and then channel decoded, inverse rate matched, deinterleaved and demodulated in known fashion (shown in Figure 3 as functional block 32').

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